IMPLEMENTATION OF SELF-NOISE SUPPRESSION TECHNIQUES FOR ULTRASONIC CORRELATION SYSTEMS

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INTRODUCTION

Pseudo-random signal correlation techniques can improve the flaw detection capability of ultrasonic NDE systems. While the correlation-based systems provide significant improvement in the signal-to-noise ratio compared to pulsed systems, their performance is limited by the so-called "self-noise" of the system. Self-noise is a result of imperfect autocorrelation characteristics of the excitation signal. Last year, we suggested some techniques for improving the flaw detection capability of continuous-mode ultrasonic NDE systems [1]. These systems use a continuously transmitted coded waveform as an excitation signal, and the received signal is processed through a correlation filter. This year, we present another new approach and demonstrate performance results and the practicability of each approach.

BACKGROUND AND THEORY

Two basic types of ultrasonic correlation systems were proposed in the past. The first type operates in the coded-pulse mode and relies on the aperiodic (linear) autocorrelation properties of the coded-pulse [2-6]. While this approach permits the use of a single transducer for transmission and reception, which makes it similar to pulse-echo methods, its performance is limited by relatively large self-noise levels. A second type of ultrasonic correlation systems, also proposed previously, operate in continuous mode and rely on the periodic (circular) autocorrelation properties of the pseudo-random waveform [7,8]. As discussed in the next section, these systems also suffer from the self-noise problem, although to a lesser extent.

The signal-processing block diagram of the ultrasonic correlation system we used is shown in Figure 1. The transmitter generates a pseudorandom excitation signal, $s(t)$, that is introduced into the test object using a suitable transmitting transducer. The scattered ultrasound, that is picked up by the receive transducer, and the additive system noise component, $n(t)$, constitute the received signal, $r(t)$. The function $h(t)$ represents the impulse response of the composite system which includes, the test object, transmit and receive transducers, and their associated electronics. The correlation filter computes the cross-correlation between the received waveform, $r(t)$ and the transmitted signal, $s(t)$. The same setup is used for both coded pulses and continuous transmission. A short review of the theory is now provided for continuity purposes.

Assuming the system to be linear and time-invariant, the output of the correlation filter can be represented by,

$$R(\tau) = \int_{t=t_o}^{(t_o+T)} s(t+\tau)\,[s(t)*h(t)+n(t)]\,dt \ . \tag{1}$$

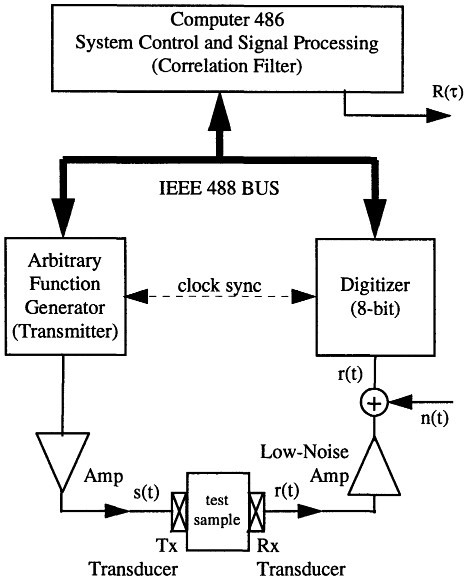

Figure 1 Signal processing block diagram for the laboratory spread-spectrum ultrasonic correlation system used for measurements.

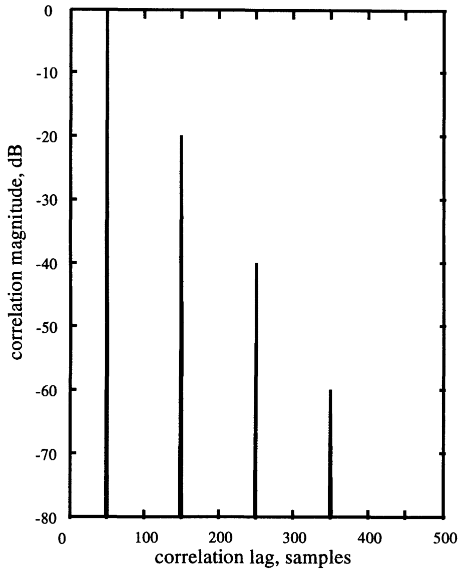

Figure 2 Idealized impulse response used for simulations.

where, T represents the integration time of the correlator. For continuous-mode systems, T is equal to the period of the transmitted signal, $s(t)$. For coded-pulse mode systems, T corresponds to the full length of the coded pulse. By interchanging convolution and integration, and computing the autocorrelation of the excitation signal, the output correlation function representing the signature of the test object is

$$R(\tau)=h(\tau)*R_{ss}(\tau)+N_r(\tau) \tag{2}$$

where $N_r(\tau)$ is the random noise component given by,

$$N_r(\tau)=\int_{t=t_o}^{(t_o+T)} s(t+\tau)\,n(t)\,dt\ . \tag{3}$$

and $R_{ss}(\tau)$ represents the autocorrelation of the excitation signal. If the autocorrelation function of the excitation waveform could be made a perfect delta function, equation (2) would become

$$R(\tau)=h(\tau)+N_r(\tau) \tag{4}$$

and we would have a estimate of the impulse response without frequency distortion and corrupted only by additive measurement noise.

In [7} and [8], the authors show that this method of estimating the impulse response, $h(t)$, provides significant improvement in SNR compared to the pulsed ultrasonic method. The focus of this paper, however, is the implementation of the "imperfect" autocorrelation function of the excitation waveform and its consequences on the estimation of $h(t)$.

SELF-NOISE IN ULTRASONIC CORRELATION SYSTEMS

There are two factors which prevent a practical correlation system from achieving the ideal results portrayed in equation (5) — the transmitted waveform is not truly random but pseudo-random and the correlation function implementation is not ideal. An ideal correlator has an infinite integration time, not the finite integration time, T, imposed by a practical correlator. For coded pulse bursts of length, T, the aperiodic autocorrelation function of the excitation waveform is given by

$$R_{ss}(\tau) = \int_{t=t_o}^{(t_o+T-\tau)} s(t+\tau)\, s(t)\, dt \ . \tag{5}$$

while for continuous-mode correlation systems, the periodic autocorrelation function is

$$R_{ss}(\tau) = \int_{t=t_o}^{(t_o+T)} s(t+\tau)\, s(t)\, dt \ . \tag{6}$$

Thus, the practical limitations imposed by (5) and (6) require that (3) be modified to

$$R_{ss}(\tau)=h(\tau)+N_s(\tau)+N_r(\tau) \tag{7}$$

where $N_s(\tau)$ represents the "self-noise" caused by the use of an excitation signal with imperfect, but practical, autocorrelation function. Self-noise is correlated with both the excitation signal, $s(t)$, and the system impulse response, $h(t)$ and its magnitude depends on the period of $s(t)$ and the nature of $h(t)$. If $h(t)$ contains a high-amplitude component caused by something such as a back wall reflection, it is very possible that the magnitude of $N_s(\tau)$ is large enough to obscure weaker components of $h(t)$.

EXAMPLES OF SELF NOISE

To predict and compare actual performance in the laboratory setup, the self-noise effects of both linear (coded pulse) and periodic (continuous) sequences in ultrasonic correlation systems were simulated. Based on using transducers with a center frequency of 5 MHZ, a pseudo-random bandpass signal, s(t), was generated from the binary phase-shift keying of a 5 MHZ (center frequency) carrier using a maximal-length - pseudo-random code of length 1023. The resulting linear (LACF, coded pulse) and periodic (PACF, continuous) autocorrelation functions (ACF) for the bandpass signal, s(t), are shown in Figures 3 and 4. Both correlation functions have their main correlation lobe at zero lag (desired response). The correlation sidelobes at non-zero lag, which are the undesired self noise, $N_s(\tau)$, are 38 dB below the main lobe for linear correlation (Fig 3) and 60 dB below the main lobe for periodic autocorrelation (Fig 4). The continuous system gives an improvement of 22 dB for a length 1023 sequence.

For comparison purposes, we simulated the effect of self noise for the two autocorrelation methods (Figures 5 and 6) for the idealized system impulse shown in Figure 2. These results show that the coded-pulse approach has a serious dynamic range limitation caused by self-noise and hence is not suitable for the most demanding applications. Its use is generally limited to the same applications where traditional pulse echo methods are used but where the material has more attenuation. The continuous-mode system has a constant self-noise level and the impulse function estimate is much better than obtained using the coded pulse. However, the self-noise level still limits it use because it obscures the smaller flaws. In the next section, we describe our approaches to reducing the self noise below both system noise and digitization noise levels.

SELF-NOISE SUPPRESSION APPROACHES

It is fairly straight forward to argue that a truly random waveform and its mathematically perfect autocorrelation function is practically unrealizable. However, there exists certain classes of deterministic but pseudorandom waveforms, which result in the complete suppression of self-noise. The last two approaches chosen for this paper are based on the design of a periodic pseudo-random waveform, $s(t)$, such that its periodic autocorrelation function (PACF) satisfies the relation,

$$R_{ss}(\tau)=0 \ \ for \ \ |\tau|>T_c \tag{8}$$

where T_c is the duration of one symbol of the maximal-length sequence and is frequently called a 'chip' interval. If N represents the length of the sequence, the period of s(t) is given by,

$$T=N{\cdot}T_c \tag{9}$$

The problem of pseudo-random waveform design for self-noise suppression directly translates into the design

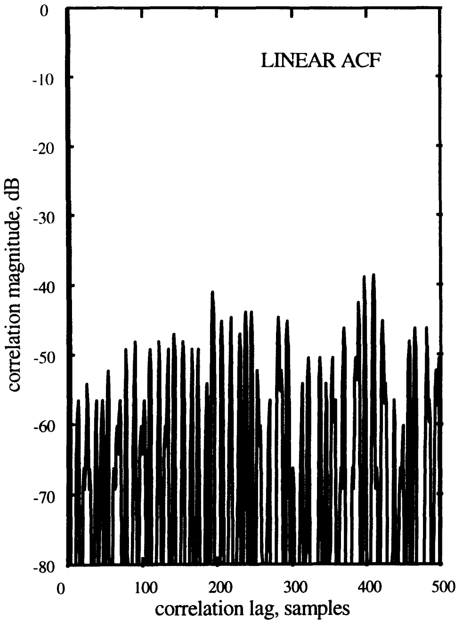

Figure 3 Linear autocorrelation function (LACF) of a bandpass excitation signal based on a maximal-length- sequence of length 1023.

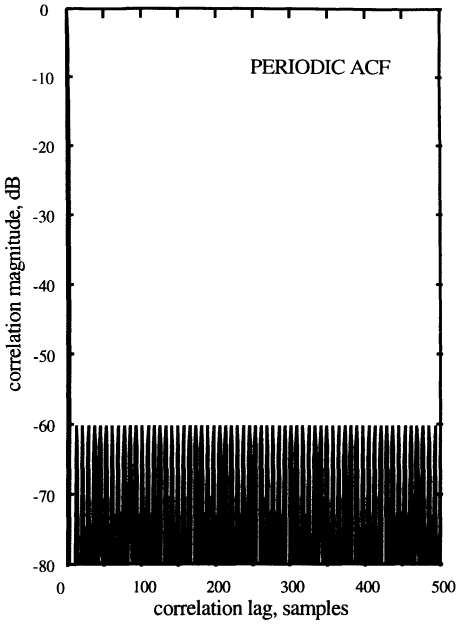

Figure 4 Periodic autocorrelation function (PACF) of a bandpass excitation signal based on a maximal-length-sequence of length 1023.

of corresponding sequences having certain desirable properties. In the following, the mathematical bases of two approaches are presented.

In our previous paper [1], we presented two approaches for obtaining perfect suppression of the self noise. In this paper, we will use the one based on perfect periodic autocorrelation defined the waveform symbol values as

$$\{z_n\} = A \pm 1 \qquad A = \frac{(-1 \pm \sqrt{M+1})}{M} \tag{10}$$

where M is the length of the sequence (1023 symbols). We refer to this as "Approach-2" – using sequences with perfect periodic autocorrelation function and amplitude offset, or simply the "amplitude offset approach." For this paper, we present a new method called "Approach-3" which is now described.

In Approach-3, we obtain a new sequence $\{z_n\}$, by making the following transformation on a maximal-length sequence $\{a_n\}$

$$\begin{aligned} z_n &= 1 & a_n &= +1 \\ z &= e^{i\phi} & a &= -1 \end{aligned} \tag{11}$$

The PACF of the resulting complex sequence is given by,

$$P_{zz}(d) = \sum_{n=1}^{M} z^*_n \cdot z_{n+d} \tag{12}$$

The product $z^*_n \cdot z_{n+d}$ can have only two values, 1 and $e^{i\phi}$. Hence equation (12) reduces to,

$$P_{zz}(d) = \alpha.1 + \beta.e^{i\phi} \tag{13}$$

Using a property of maximal-length sequences, it can be shown that,

$$P_{zz}(d \neq 0) = \frac{(M-1)}{2} + \frac{(M+1)e^{i\phi}}{2} \tag{14}$$

The constraint equation that will set the magnitude of $P_{zz}(d\neq 0)$ equal to zero is

$$(M-1)+(M+1).\cos(\phi)=0 \tag{15}$$

which gives the value of the phase angle as,

$$\phi=\cos^{-1}(-\frac{M-1}{M+1}) \tag{16}$$

Hence, if ϕ is chosen according to the above relation, the sequence $\{z_n\}$ will be based upon two elements '+1' & $e^{i\phi}$, and the PACF of the sequence will be perfect. This means that,

$$P_{zz}(d\neq 0)=\sum_{n=1}^{M} z_n z*_{n+d} = 0 \tag{17}$$

$$P_{zz}(d=0)=\sum_{n=1}^{M} z_n z*_{n} = M \tag{18}$$

Approach-3 will be referred to as the "phase-offset method."

For comparison purposes, it is necessary to summarize our original method (Approach-1) to generating the random continuous waveform. It was based on a well-known method of carrier modulation called "binary phase shift keying" (BPSK) where $s(t)$ is generated using the following transformations for sequence values to phases to signal:

$$\begin{aligned} z_n &= +1 \rightarrow \theta(t)=1 \\ z_n &= -1 \rightarrow \theta(t)=\pi/2 \qquad s(t) = A\cos[2\pi f_0 t + \theta(t)] \end{aligned} \tag{19}$$

Although the original method had low self-noise, it was not zero. The new method (phase-offset) defined in the same format is

$$\begin{aligned} z_n &= +1 \rightarrow \theta(t)=1 \\ z_n &= -1 \rightarrow \theta(t)=\cos^{-1}(-\frac{M-1}{M+1}) \qquad s(t) = A\cos[2\pi f_0 t + \theta(t)] \end{aligned} \tag{20}$$

and we remind the reader that the waveform values vary as a function of M, the code length.

SYSTEM IMPLEMENTATION AND SIMULATION

The new phase-offset method was implemented on a general purpose lab-grade instrument (Fig. 1) and its performance evaluated by making measurements on a test sample. A 486PC is connected to the arbitrary function generator (AFG) and to the 8-bit digitizer through the IEEE-488 interface. The excitation waveform, $s(t)$, was generated in the PC and downloaded to the AFG. The waveform was based on a 10th order maximal length sequence - giving sequence length of 1023. The received signal, after amplification, was digitized at a 100 MHZ sample rate using an eight bit digitizer. The correlator was implemented with digital signal processing algorithms on the PC.

It is important to understand the instrument limitations before performing the experiment and interpreting the results. Of particular consequence is the effect of quantization noise generated as a result of the digitization of the received waveform. The quantization noise produces a random noise component in the measured signature function, as represented in (4) and (7). For an 8-bit quantizer, the quantization noise is on the order of -40 dB. The test object was a circular disk of plastic material, 1/4 inch thick and ½ inch in diameter. The measured impulse response was expected to consist of a through-transmission component followed by regularly spaced and attenuated components corresponding to the multiple reflections from the two flat surfaces of the disk.

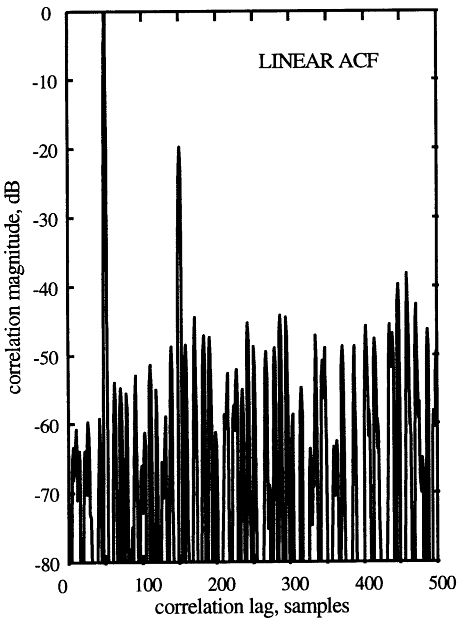

Figure 5 Impulse response simulation using coded-pulse correlation system.

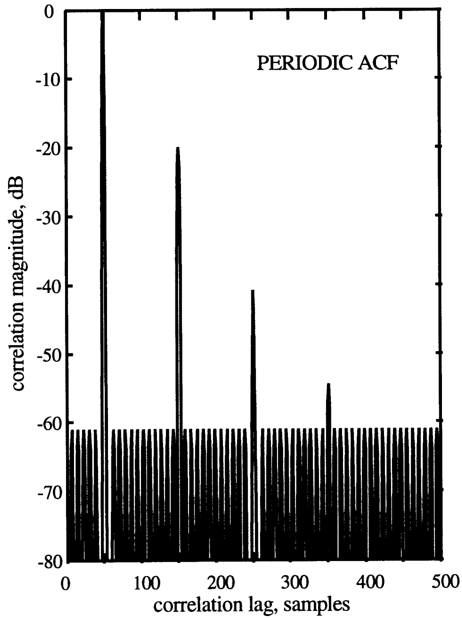

Figure 6 Impulse response simulation using continuous-mode correlation system.

The measured impulse response using our previous method (BPSK, Approach-1) is shown in Figure 7. It has an almost constant self-noise level of -60 dB, which matches well with the theoretical results. The self-noise limits the detectability limit for the BPSK approach and , in this example, reflections beyond the 6th cannot be observed. Any signal component weaker than -60 dB is obscured by the self noise and is undetectable. Figure 8 shows the equivalent test result using amplitude offset (Approach-2). A casual comparison might only reveal that there is an improvement in the detectability limit of about 20 dB since the noise floor has gone down to between -60 and -80 dB. Thus making it possible to "see" the 7-th through the 10-th multiple reflections which were previously obscured. However, a detailed analysis of the result shows that the detectability limit is now determined by the random noise level (instead of the self-noise level) and the noise floor in this figure corresponds to the random noise component of (4). Finally, the equivalent results for the new offset phase method (Approach-3) is shown in Figure 9.

PRACTICAL CONSIDERATIONS

Approach-3 produces a signal which is also described as offset biphase modulation since the difference between the two phase angles is not exactly 180 degrees. Also, the waveform has a constant envelope so amplitude distortions are not a big concern. By using new digital techniques, such as arbitrary waveform generation, offset BPSK can be generated very conveniently. Previously, generation of an accurate offset biphase modulated signal using analog circuits was very difficult. Finally, this optimum waveform design is derived only for maximal-length sequences. This is not a serious limitation since most pseudo-random correlation systems are designed with maximal-length sequences because of their well-known and effective randomness properties. Perhaps the only limitation is that the maximal-length sequences exist only in lengths $(2^n - 1)$, where n is a positive integer. The presented techniques for self-noise suppression can, however, be extended to a much wider class of pseudo-random sequences.

CONCLUSION

By employing a carefully designed, periodic pseudo-random excitation waveform and a periodic correlation filter, it is possible to completely eliminate the self-noise in ultrasonic correlation systems. The remaining limits then become random system noise and digitization noise. Performance measurements in the laboratory indicate that the new method is very effective.

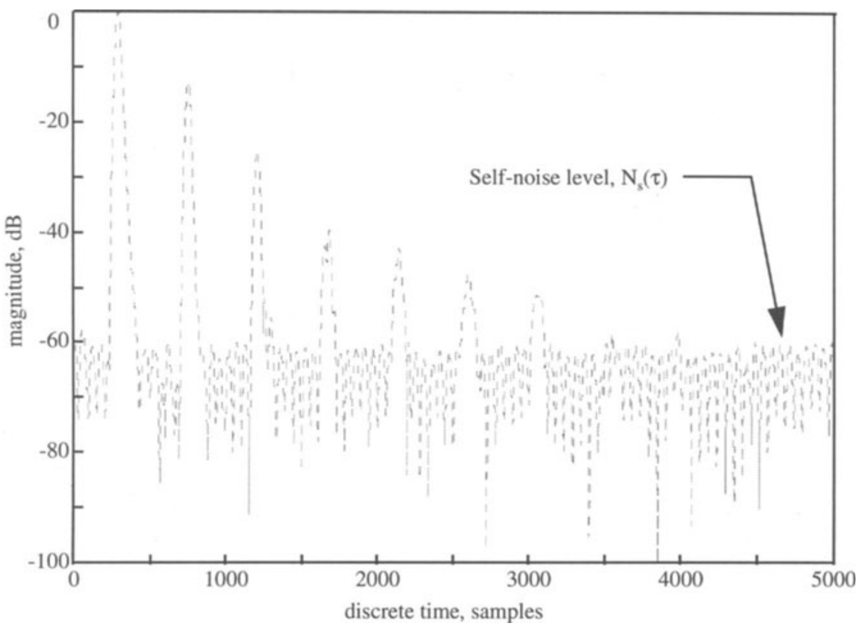

Figure 7 Impulse response (correlation signature) of plastic disk using the previous BPSK waveform (Approach-1).

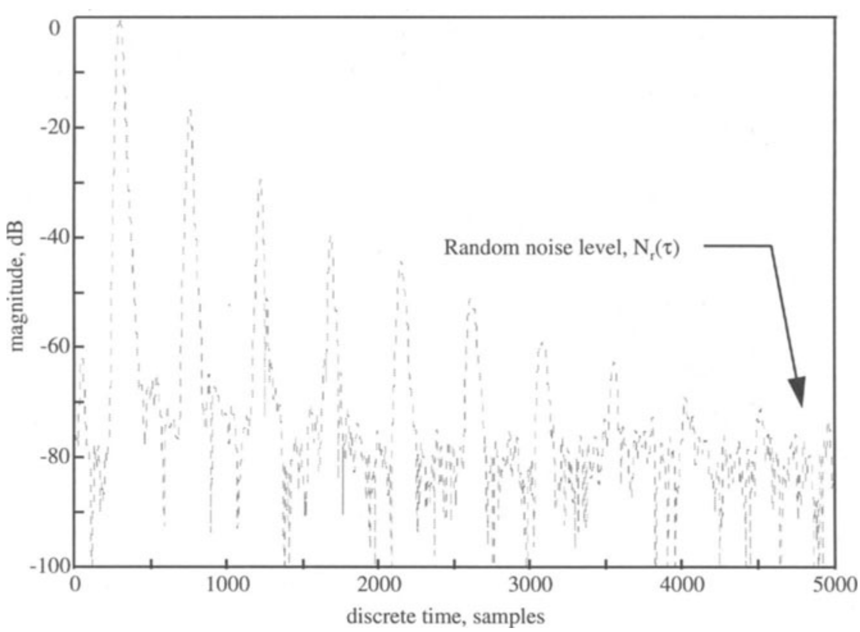

Figure 8 Impulse response (correlation signature) of plastic disk using sequences with perfect periodic autocorrelation function and amplitude offset (Approach-2).

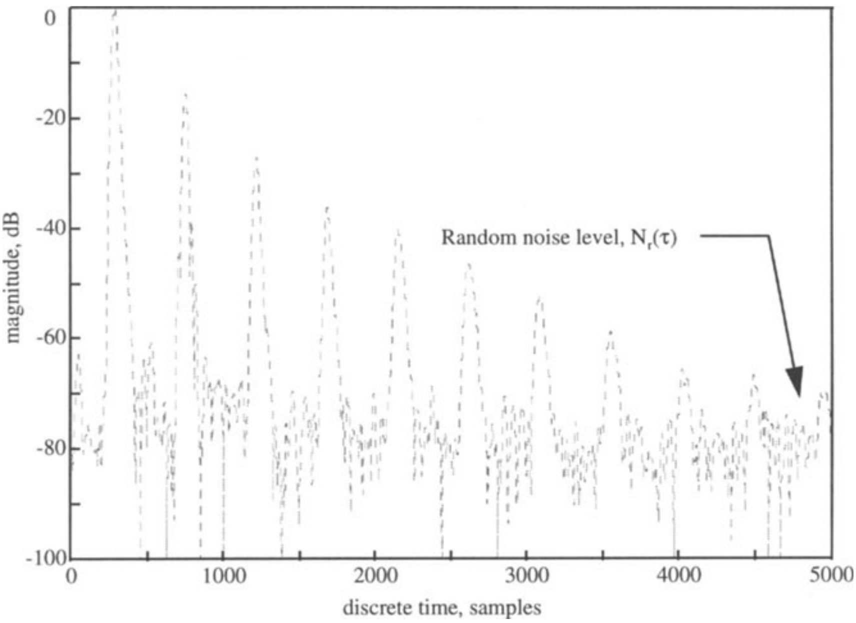

Figure 9 Impulse response (correlation signature) of plastic disk using sequences with perfect periodic autocorrelation function and phase offset (Approach-3).

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